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All-tap Fractionally Spaced, Serial Rake Combiner Apparatus  
and Method

Field of the Invention

The present invention is related to the reduction of multi-path interference in direct sequence spread spectrum (DSSS) signals whereby a primary signal is received along with a plurality of delayed, or reflected, signals. The received symbols may form a codeword, and in the case of spread spectrum IEEE 802.11b 1Mb and 2Mb data rates, the codeword may be a Barker codeword for which a decision needs to be made as to whether the codeword for a one (1) or a negative one (-1) was received. In wireless communications systems, the received signal comprises the signal traveling the shortest path accompanied by a plurality of reflected, phase shifted, and later-arriving copies of the original signal. The invention is more specifically directed to spread spectrum wireless communications links such as IEEE

1 802.11b which operate in the presence of multi-path  
2 reflections.

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## 5 Background of the Invention

6 Prior art wireless communication systems are defined in  
7 IEEE protocols for 802.11b. IEEE 802.11b is an example of a  
8 spread spectrum wireless protocol whereby a transmitter  
9 emits a typically isotropic or multidirectional transmission  
10 of a sequence of baseband codewords translated in frequency  
11 to a carrier frequency, where the codewords are phase  
12 modulated according to a known sequence, which is referred  
13 to as a chipping code. For 1Mb and 2Mb IEEE 802.11b data  
14 rates, the chipping code is a Barker code, which has the  
15 properties of maximum self-correlation and minimum cross-  
16 correlation. When a typical IEEE 802.11b receiver is placed  
17 within a reception distance from the source, it is common  
18 for a plurality of signals to arrive at the receiver. These  
19 signals may consist of a first signal which travels a  
20 shortest path, followed by a plurality of reflected signals  
21 which have path lengths longer than the shortest path.  
22 Occasionally, the shortest path will be through an  
23 attenuating medium, such that the shortest path does not  
24 correspond to the strongest signal. The receiver then must  
25 recover this plurality of signals and apply some method of

1 filtering or reconstruction to regenerate the transmitted  
2 codeword. One prior art method uses channel sounding,  
3 whereby the communications channel is characterized by  
4 sending a known signal such as a preamble. Then applying an  
5 inverse filter function compensates the channel for the  
6 reflections, thereby restoring the channel. The process  
7 for generating the inverse filter coefficients such as in a  
8 finite impulse response (FIR) filter typically involves  
9 identifying the plurality of received signals, applying a  
10 delay and a weight to each signal, and performing the signal  
11 reconstruction with an FIR filter. Since the coefficients  
12 of an FIR filter must be computed from the delays and  
13 weights, this may require extensive computation complexity  
14 and computation time, and may not be achievable during the  
15 short preamble time of the incoming packet, where the  
16 coefficients must be computed before the packet data  
17 arrives.

18 Figure 1 shows a wireless Local Area Network (LAN)  
19 operating in a typical office environment, wherein two  
20 mobile stations may be transmitting and receiving data  
21 through the wireless medium. While both station 100 and  
22 station 102 are typically transmitting and receiving  
23 wireless signals, for simplicity of analysis, station 100 is  
24 shown as a transmitter and station 102 is shown as a  
25 receiver. In the figure, receiver 102 receives a first

1 signal from an attenuated path R1 112 r1 114 with an  
2 attenuating barrier 106, followed by the next signal with  
3 the path length R2 108 and reflection r2 220, followed by  
4 the next signal with the path length R3 116 and reflection  
5 r3 118. In this figure, a small number of such path  
6 reflections are shown, however as a practical matter, an  
7 infinite number of such reflection paths exist, shown  
8 generally by path  $R_n$  120 and reflected path  $r_n$  222. In the  
9 wireless IEEE 802.11b protocol, the use of spread spectrum  
10 transmission and reception techniques improves the bit error  
11 rate (BER) in the presence of multi-path reflections to the  
12 receiver. However, the BER can be greatly reduced if the  
13 various reflections ( $r_1$  to  $r_n$ ) are used to make a decision  
14 at the receiver demodulator, rather than using the strongest  
15 (relatively) signal only, ignoring all other multi-path  
16 reflections present. Moreover, the absence of a direct path  
17 in such communications (as shown by path R1-r1) means that  
18 the information is highly dispersed and arrives strongly  
19 attenuated and through various path lengths so that the  
20 single correlation peak energy is spread out over many  
21 correlation peaks arriving at later times. A temporal  
22 Maximal Ratio Combiner (MRC) operating on the spread  
23 spectrum correlation peaks would give a much higher BER than  
24 a simple decision based on a single maximum peak. The  
25 combining of signal energy arriving at different times

1 associated with the transmission of data is the basic  
2 principle on which the rake combiners for spread spectrum  
3 receivers operate. A plurality of weighted delays in the  
4 form of a Feed Forward Filter (FFF) is used to cause the  
5 various correlation peaks associated with the delayed  
6 arrival of a plurality of multipath reflections of a given  
7 codeword to recombine into a single peak. Additionally,  
8 since the multipath reflections cause the rotation of  
9 incoming codeword symbols out of their original quadrature  
10 channels, an additional function of the MRC is to apply  
11 phase compensation to rotate these symbols back in to the  
12 original quadrature channels, so that received signal energy  
13 is rotated back into the proper I and Q channels, thereby  
14 adding to the signal that is received on the direct path.

15 Figure 2 shows a decision block for a direct sequence  
16 spread spectrum receiver 140 that includes a sliding  
17 correlator 130 which correlates the Barker codeword 131 with  
18 the signed incoming quadrature symbol stream 142. The  
19 incoming quadrature symbol stream 142 comprises an analytic  
20 signal formed by receiver quadrature mixing using the  
21 relationship  $e^{j\phi(t)} = \cos(\phi(t)) + j*\sin(\phi(t))$ , where the  
22 first term represents in-phase (I) information in the real  
23 channel and the second term represents quadrature (Q) symbol  
24 information in the imaginary channel. Each channel of the  
25 incoming symbol stream is digitized using an Analog to

Digital Converter into positive and negative values N-bit wide (where N is the width of the converter outputs), where the positive and negative values of the input stream contain Barker codewords:

I channel 1={+1,-1,+1,+1,-1,+1,+1,+1,-1,-1,-1};

0={-1,+1,-1,-1,+1,-1,-1,-1,+1,+1,+1}

Q channel 1={+j,-j,+j,+j,-j,+j,+j,+j,-j,-j,-j};

0={-j,+j,-j,-j,+j,-j,-j,-j,+j,+j,+j},

where the Q channel codewords are simply the I channel codewords multiplied by j. A phase rotator function in the aforementioned receiver digital signal processing causes the symbols to be rotated into the I or Q channels of the correlator input 142.

For an 11 bit Barker code, the soft (N-bit wide) I and Q channels are cross-correlated with the respective Barker codewords for 1. This cross-correlation of the input stream containing Barker codes for 1 produces the impulsive value +11, while the cross correlation of the input stream containing 0 produces the impulsive value -11. In this manner, the cross-correlation of the incoming data against the Barker codewords produces I and Q values which have a peak impulsive positive response for the 1 codeword, a peak impulsive negative response for the 0 codeword, and approximately 0 at all other times. Because of the impulsive nature of the cross correlation response, the

1 incoming quadrature signal 142 is typically presented as an  
2 over-sampled signal at twice the bit rate, or 22Mhz, so that  
3 information for compensating timing drift between the  
4 transmit and receive DAC and ADC clocks is retained in the  
5 digitized signal. Additionally as mentioned above, the  
6 incoming quadrature symbol stream is multi-valued rather  
7 than binary, so the correlation outputs 138 and 139 are an  
8 over-sampled, multi-valued quadrature data stream. The code  
9 acquisition function 132 maintains a correlation output  
10 history and determines the peak correlation points, which  
11 are used to produce a synchronization pulse accompanied by  
12 an indication for the start and end of the codeword window.  
13 Once codeword acquisition is achieved, the baseband Direct  
14 Sequence Spread Spectrum (DSSS) codewords 142 are presented  
15 to a simpler fixed window correlator 134, which performs a  
16 correlation 137 of the incoming quadrature symbol stream 142  
17 to the Barker codeword 143. The multi-valued quadrature  
18 result of this cross correlation output 139 is stored in  
19 memory 135 at a synchronization time determined by code  
20 acquisition function 132, which presents a pulse at the  
21 point of maximum positive or negative signal, as described  
22 earlier. Decision block 136 examines the positive or  
23 negative value stored in memory 135, and decides whether a  
24 Barker codeword for 1 or 0 was received. For a 2Mb data  
25 rate, the fixed window correlator 134 is separately

1 correlating the I and Q streams with the Barker codeword,  
2 thereby storing in memory 135 1 bit of decision data from  
3 the I and Q stream for each 1 us codeword interval. For a  
4 1Mbps data rate, the phase encoding is in the I (or Q)  
5 channel alone, and the Q (or I) channel contains only  
6 correlation and thermal noise. For the 2Mbps data rate, the  
7 number of possibilities double since now the receiver can  
8 get (1,0), (-1,0), (0,1) or (0,-1) in each 1us duration,  
9 i.e., 2-bits of information every microsecond. The spread  
10 spectrum signal comprises one of the two 11 bit Barker  
11 codewords occurring within 1 us boundaries for IEEE 802.11b  
12 wireless communication systems. Examining the codeword  
13 output 138, the code acquisition function 132 of figure 2  
14 stores several windows of samples and extracts a  
15 synchronization signal 144. This synchronization pulse and  
16 a window signal (shown as timing 151 in figure 3) to  
17 establish the framing of codewords is sent to the fixed  
18 window correlator 134 and to the decision function 136  
19 which determines which chip code is the closest  
20 approximation. In direct sequence spread spectrum (DSSS)  
21 signals, chip codes that are dependant on the particular  
22 802.11b data rates are used for encoding and decoding.  
23 These chip codes are Differential Binary Phase Shift Keying  
24 ((D)BPSK) using Barker codes for 1Mbps, and Differential  
25 Quadrature Phase Shift Keying ((D)QPSK) using Barker codes



1 for 2Mbps. The higher data rates of 5.5Mb and 11Mb use  
2 Complementary Code Keying (CCK). In the present invention,  
3 we are concerned with 1Mbps and 2Mbps data rates using 11  
4 bit Barker codes. It is essential that the decision block  
5 for a direct sequence spread spectrum receiver 140 of figure  
6 2 properly synchronize to the peak and window of the 11 bit  
7 codewords, done during the preamble interval of the packet  
8 with sliding correlator 130. An unsynchronized stream of  
9 incoming codewords 142 is referred to as a symbol stream,  
10 and once the synchronization to the 11 bit codewords is  
11 accomplished, this is referred to as a codeword stream. The  
12 codeword stream is then converted into binary values  
13 comprising a data stream 148. Examining the quadrature  
14 multi-valued output of the sliding correlator 130, a datum  
15 point may be defined as the strongest correlation response  
16 of the present symbol. Pre-cursors are signals received  
17 prior to the datum point, and post-cursors are signals  
18 received after the datum point. The sample window includes  
19 pre-cursors, the datum point, post-cursors, and has a  
20 duration equal to a symbol length. Directly received signal  
21 in the presence of attenuation followed by multi-path  
22 reflection produces correlation output 138 which comprises  
23 pre-cursors, followed by the strongest correlation response  
24 of the present symbol earlier referred to as a datum point,  
25 and finally post-cursors representing the multipath

1 reflections. In combination, these signals represent the  
2 total signal power available for use by the decision  
3 processor. Although received at different times, the sum of  
4 these signals in the channel represent the total signal  
5 power available. In one prior art receiver, the strongest  
6 correlation response is selected for synchronization signal  
7 144 and the pre-cursors and post-cursors are ignored, which  
8 results in reduced signal to noise, since the noise level is  
9 constant while the signal power associated with the pre-  
10 cursor symbols and post-cursor symbols is discarded. It is  
11 desired to use a plurality of symbols including pre-cursors  
12 and post-cursors to generate a composite signal which uses  
13 the pre-cursors and post-cursors so that the signal to noise  
14 ratio is improved.

15         Figure 3 shows waveform 150 of output 138 of the  
16 sliding correlator 130, both from figure 2, for a Barker  
17 codeword 1 followed by a Barker codeword 0 presented to the  
18 input 142 from figure 2. In the absence of multi-path  
19 reflection or noise, the sliding correlator 130 output 138  
20 would appear as shown in curve 150 of figure 3. Every  
21 sample time, the incoming symbol stream 142 of Barker codes  
22 for 1 followed by 0 is correlated 133 with Barker codeword  
23 131 in the sliding correlator 130, which produces the  
24 positive peak 152a followed by a negative peak 152b. These  
25 peaks are detected by code acquisition function 132, which

1 uses an algorithm for defining a window comprising the  
2 timing points 0 through 10 151 of figure 3, as shown. Once  
3 this synchronization is achieved, the timing of the  
4 remainder of the Barker codewords may be maintained by the  
5 precision of the clock oscillators of the receiver and  
6 transmitter. The synchronization signal 144 is applied to  
7 fixed window correlator 134, which selects one of two  
8 codewords in the I (or Q) channel for a 1Mb data rate, or  
9 one of four codeword combinations in the I and Q channel for  
10 a 2Mb data rate. In the absence of noise or multi-path  
11 reflections, the decision block 140 for a direct sequence  
12 spread spectrum receiver 140 of figure 2 performs well, and  
13 the cross correlation of the Barker codeword 131 with the  
14 incoming symbol stream produces the correlator output 150 as  
15 shown in figure 3. In the presence of multi-path  
16 reflections, degradation of the cross correlation output 138  
17 occurs. For example, given the signal paths shown in figure  
18 1, an attenuated signal traveling path r1 114 followed by a  
19 stronger reflected signal r2 220, followed by a weaker  
20 reflected signal r3 118 would produce signal 154 comprising  
21 the attenuated response r1 156a, reflected and phase shifted  
22 maximum signal r2 158a, and reflected and attenuated signal  
23 r3 160a respectively as shown in figure 4. The phase  
24 rotator (not shown) which performs phase rotation of the  
25 incoming signal stream prior to delivery as stream 142 uses

1 maximum level signals as the datum for this phase rotation.  
2 In figure 4, the phase rotation causes maximum responses  
3 158a and 158b to be rotated in phase until they are aligned  
4 in the I channel. Using maximum response only, the code  
5 acquisition function 132 could incorrectly determine that r2  
6 158a is code position 0 based on purely amplitude  
7 considerations, when the actual synchronization time is r1,  
8 as shown by the alignment with the time interval 0 155.  
9 Figure 4 shows the I and Q components of the incoming  
10 signal, which have been phase rotated by the phase shifter  
11 (not shown), such that the strongest signal 158a has been  
12 shifted exclusively into the I channel, and precursor 156a  
13 is shifted in phase, as is postcursor reflection 160a , as  
14 evidenced by the shifting of energy into the Q channel for  
15 156a and 160a. Similarly, for the 0 symbol, strongest  
16 signal 158b is only seen in the I channel, while precursor  
17 156b and postcursor reflection 160b is shifted in phase.

18 Figure 5 shows a prior art rake combiner 190 with the  
19 rake having three fingers. This allows for improved signal  
20 to noise performance and multi-path reflection reduction,  
21 and implements a function known as temporal maximal ratio  
22 combining (MRC). An MRC identifies a finite number of  
23 strongest signals, and combines them to rearrange the signal  
24 power of each of the delayed signals into a single signal  
25 with maximum power such that there is maximum likelihood of

1 decoding the signal correctly. In figure 5, the baseband  
2 signal stream 194 is comprised of the figure 4 signals  $r_1$   
3 156a,  $r_2$  158a occurring  $\tau_1$  later, and  $r_3$  160a occurring  $\tau_2$   
4 thereafter. The code acquisition 162 generates a  
5 synchronization pulse 196 from the output of the sliding  
6 correlator 192, similar to sliding correlator 130 of figure  
7 2, but delayed by  $\tau_1 + \tau_2$ . The baseband symbol stream 194 is  
8 provided to correlator 168, and the same signal is provided  
9 to correlator 166 delayed by  $\tau_2$ , and to correlator 164  
10 delayed by  $\tau_1 + \tau_2$ . In this manner, at the instant of  
11 synchronization pulse 196, correlator 164 is providing  
12 signal associated with  $r_1$  in figure 4, correlator 166 is  
13 providing signal associated with  $r_2$  in figure 4, and  
14 correlator 168 is providing signal associated with  $r_3$  in  
15 figure 4. These signals are multiplied by a conjugated  
16 series of complex coefficients  $\alpha_0$ ,  $\alpha_1$ , and  $\alpha_2$ , respectively.  
17 The conjugated complex coefficients  $\alpha_0$  174,  $\alpha_1$  176, and  $\alpha_2$   
18 178 are a linear array of complex numbers with a length  
19 equal to the length of the incoming signal stream window,  
20 repeated continuously, as it known to one skilled in the art  
21 of finite impulse response filters operating on periodic  
22 signals. The correlators 164, 166 and 168 are multiplied by  
23 the complex conjugate of the linear array of conjugated  
24 complex coefficients  $\alpha_0$  174,  $\alpha_1$  176, and  $\alpha_2$  178,  
25 respectively. The conjugated complex coefficients  $\alpha_0$  174,

1  $\alpha_1$  176, and  $\alpha_2$  178 reverses the sign of the imaginary part of  
2 the coefficient. For example, the complex conjugate of  $(a +$   
3  $jb)$  is  $(a - jb)$ , and vice-versa. Multiplying a value by its  
4 complex conjugate has the effect of converting all of the  
5 signals in the imaginary channel into the real channel.  
6 The complex multi-valued outputs of multipliers 180, 182,  
7 and 184 are summed respectively by adder 186, and provided  
8 to decision block 188, which performs in the same manner as  
9 decision block 136 of figure 2.

10 Figure 6 shows a maximal ratio combiner processor 201,  
11 where the signals and later reflections are delayed and  
12 multiplied by a linear array of conjugated complex  
13 coefficients  $\alpha_0$  208,  $\alpha_1$  210, and  $\alpha_2$  212, then summed to form  
14 a signal containing maximum combined signal power.  
15 Similarly to the rake combining signal processing system of  
16 figure 5, the function of the MRC processor 201 is to delay  
17 pre-cursor signal 156a of figure 4 with delay element 204  
18 and to delay the strongest symbol 158a of figure 4 with  
19 delay element 206 such that these signals align in time with  
20 post-cursor 160b of figure 4. The MRC processor 201 then  
21 multiplies each stream of pre-cursors, the strongest signal,  
22 and post-cursors separately with a linear array of  
23 conjugated complex coefficients  $\alpha_0$  208,  $\alpha_1$  210, and  $\alpha_2$  212,  
24 respectively, and then sums the aligned and multiplied  
25 streams in adder 220, thereby producing reflection-

1 compensated output 228. This reflection-compensated output  
2 228 is then fed to correlator 222, which is gated by  
3 synchronization pulse 226 similar to synchronization pulse  
4 196 in figure 5. The effect of multiplying by the  
5 conjugated complex coefficients  $\alpha_0$  208,  $\alpha_1$  210, and  $\alpha_2$  212  
6 is the phase rotation of reflected multi-path signals back  
7 into the proper channel. For both of the maximal ratio  
8 combiners in figures 5 and 6, the computation of delay  
9 coefficients for delay elements 170 and 172 of figure 5,  
10 delay elements 204 and 206 of figure 6, as well as the  
11 linear arrays of conjugated complex coefficients  $\alpha_0$  174,  $\alpha_1$   
12 176, and  $\alpha_2$  178 in figure 5, and the linear arrays of  
13 conjugated complex coefficients  $\alpha_0$  208,  $\alpha_1$  210, and  $\alpha_2$  212  
14 in figure 6, are done based on the strongest correlation  
15 signals. These conjugated complex coefficients must be  
16 computed based on the amplitude and the phase of the  
17 particular pre-cursors and post-cursors present, which  
18 depends upon the nature of multipath present.

19 Figure 7 shows the waveform for time-shifted cross  
20 correlation (normalized so that peak is one) for the 256  
21 possible complementary code keying (CCK) codewords used in  
22 the 11Mbps data rate of IEEE 802.11b wireless communication  
23 systems. It can be seen that the cross-correlation response  
24 (with time shifts of  $\pm 7$  symbols) of up to 0.4 times the  
25 peak (autocorrelation) rules out any major multipath

1 tolerance of a system using the autocorrelation properties  
2 of CCK codewords to do an MRC combining. This has the  
3 additional implication that the maximal ratio combiner 201  
4 of figure 6 could be placed in the signal processing stream  
5 for both the Barker and CCK codewords. This would not be  
6 possible with the rake combining signal processing system  
7 190 of figure 5, since the multiplication and summing is  
8 done post-correlation. But figure 7 shows the CCK codeword  
9 response for cross-correlation is poor and hence any system  
10 relying on such an MRC combining for multipath tolerance  
11 would fail for the CCK rates under multipath.

12 Figure 8 shows a waveform for the sliding correlation  
13 output for Barker codewords. The figure shows the 1 us  
14 timing and response of a noise-free signal presented to a  
15 sliding correlator operating at twice the bit rate, or  
16 22Mhz. Figure 9 shows the waveform for the sliding  
17 correlation output for noise input - noise that would  
18 ordinarily be present with the signal of figure 8.

19

## 20 Objects of the Invention

21 A first object of the invention is a system for maximum  
22 ratio combining whereby a sliding correlator provides window  
23 and synchronization information which is used to generate a  
24 training decision signal which is then used to invert the



1 received codewords for 0 and not invert codewords for 1, and  
2 to sum these with the contents of a channel profile memory.

3 A second object of the invention is to use the values  
4 stored in a channel profile memory by synchronizing them to  
5 the incoming stream of correlated signals which are stored  
6 in the channel profile memory, then conjugated and  
7 multiplied with the incoming correlation stream to produce a  
8 first output, which is accumulated and added to produce a  
9 rake combined output during a window. The rake combined  
10 output thereby includes signal power from all post-cursors  
11 and pre-cursors, thereafter using the accumulated output to  
12 make a decision on which value was received.

13 A third object of the invention is the generation of  
14 channel profile values from a stream of correlation peaks,  
15 each bounded by a window, whereby a channel profile memory  
16 having a length equal to said window is initialized,  
17 thereafter adding to the channel profile memory each stream  
18 of incoming correlation peaks during said window, the  
19 accumulation of data into the channel profile memory known  
20 as a training time, which is followed by a decision time,  
21 whereby said stream of correlation peaks is synchronized  
22 with said channel profile RAM, and the correlation peaks are  
23 multiplied by a complex conjugate of the contents of the  
24 channel profile memory, thereby accumulating the sum of all  
25 such multiplications during said window, and making a

1 decision on the value of the symbol at the end of each said  
2 window.

3

#### 4 Summary of the Invention

5       During a training interval, a stream of baseband  
6 symbols from a wireless spread spectrum receiver is  
7 correlated with one or more codeword symbols into positive  
8 and negative correlation peaks. These correlation peaks are  
9 averaged into a channel profile Random Access Memory (RAM),  
10 and also generate a periodic window signal, during which  
11 window the pre-cursors and post-cursors for the present  
12 symbol are contained. After the training interval, a  
13 decision interval results in the contents of the channel  
14 profile RAM being conjugated and multiplied with the  
15 correlation peaks, producing a multiplier output. The  
16 multiplier output is fed to an accumulator, which  
17 initializes at the beginning of the window, and then adds  
18 each subsequent multiplier output during the window period,  
19 and finally is sampled at the end of the period to produce a  
20 decision output. This decision output includes the signal  
21 power from all the pre-cursors and post-cursors of the  
22 present codeword, and thereby provides a stream of  
23 decisions. The window interval may be chosen to allow  
24 approximately 20% precursors compared to the largest  
25 correlation signal, and 80% post-cursors, or any ratio of

1 precursors and post-cursors which characterizes the  
2 operating condition of the receiver.

3

#### 4 Brief Description of the Drawings

5 Figure 1 shows a transmitter coupled to a receiver  
6 through a plurality of reception paths.

7 Figure 2 shows a prior art decision block for a direct  
8 sequence spread spectrum receiver.

9 Figure 3 shows a waveform for a sliding correlator  
10 without multi-path reflections.

11 Figure 4 shows the waveform for a sliding correlator  
12 with multi-path reflections.

13 Figure 5 shows the block diagram for rake combining  
14 signal processing system, a type of maximal ratio combiner,  
15 with a plurality of fixed-window correlators.

16 Figure 6 shows the block diagram for a maximum ratio  
17 combiner as an FIR filter before a single fixed window  
18 correlator.

19 Figure 7 shows the waveform for the normalized, time-  
20 shifted cross correlation output for complementary code  
21 keying (CCK) codewords.

22 Figure 8 shows the waveform for the sliding correlation  
23 output for Barker codewords.

24 Figure 9 shows the waveform for the sliding correlation  
25 output for noise input.

1        Figure 10 shows the block diagram for an all-tap,  
2        fractionally-spaced serial rake combiner processor.

3        Figure 10a shows an alternate function for the  
4        demultiplexer of figure 10.

5        Figure 11a shows the waveform for Barker chip codes for  
6        an 00010 data pattern.

7        Figure 11b shows the I output of a sliding correlator  
8        operating at 11Mhz when presented with the data of figure  
9        11a.

10       Figure 11c shows the Q output of a sliding correlator  
11       operating at 11Mhz when presented with the data of figure  
12       11a.

13       Figure 11d shows the I output of a sliding correlator  
14       operating at 22Mh when presented with the data of figure  
15       11a.

16       Figure 11e shows the Q output of a sliding correlator  
17       operating at 22Mhz when presented with the data of figure  
18       11a.

19       Figure 11f show the I output of a sliding correlator  
20       operating at 22Mhz when presented with multi-path  
21       reflections of the data of figure 11a.

22       Figure 11g show the Q output of a sliding correlator  
23       operating at 22Mhz when presented with multi-path  
24       reflections of the data of figure 11a.

1        Figures 12a and 12b show the waveforms for sliding  
2 correlator I and Q outputs with multi-path reflections.

3        Figure 12c shows the output of the training decision  
4 block during the training interval.

5        Figures 13a and 13b show the I and Q channel memory  
6 contents during the training period.

7        Figures 14a and 14b show the waveforms for the  
8 operation of the all-tap fractionally spaced serial combiner  
9 of figure 10.

10  
11  
12        Detailed Description of the Invention

13        Figure 10 shows an all-tap, fractionally-spaced serial  
14 rake combiner processor 284. Its operation is best  
15 understood in combination with the waveforms of figures 11a  
16 through 11g. As with the maximal ratio combiners of figures  
17 5 and 6, a stream of incoming quadrature baseband symbols  
18 286 comprising multi-valued I channel and a Q channel from  
19 the output of the baseband conversion radio frequency (RF)  
20 section (not shown) are provided to sliding correlator 250.  
21 Sliding correlator 250 then performs a complex cross-  
22 correlation 251 of the stream of incoming quadrature  
23 baseband symbols 286 with the I and Q Barker codewords 253  
24 to produce a stream of correlation peaks 258. During a  
25 training interval which may occur during the preamble time

1 of the received packet, these correlation peaks 258 are fed  
2 to a training decision function 252. This training decision  
3 function 252 then examines the previous samples of  
4 correlation peaks 258 by using a decision processor 285 to  
5 generate a window signal 288. This window signal 288  
6 indicates the extent and framing of each codeword and  
7 decision signal, as well as a training decision output 256  
8 which is a decision made on each codeword which is used  
9 during the training interval. The decision processor 285  
10 may formulate a decision on the current codeword and  
11 codeword window by selecting the largest peak, and then  
12 framing the window function to allow for 20% precursors and  
13 80% post-cursors about the largest peak. This may be  
14 similarly done with any other window function which includes  
15 pre-cursors and post-cursors. Decision processor 285 may use  
16 any other method known in the art of framing and  
17 synchronizing a multi-path received signal. Storage element  
18 293 used in conjunction with decision processor 285 may rely  
19 on a single maximum value as a datum point for windowing  
20 pre-cursors and post-cursors, or it may store the entire  
21 received window of correlation output data for data analysis  
22 to determine the window signal. Figure 11a shows the Barker  
23 codeword for 0 of 01001000111, and the Barker codeword for 1  
24 of 10110111000, and the serial stream for data word 00010.  
25 The Barker codewords have the property of producing -1 when

1 auto-correlated except for location 0 in autocorrelation,  
2 where they produces a maximum output of 11. Additionally,  
3 in a typical system, the individual 1s and 0s of each Barker  
4 codeword are received as amplitude values which include  
5 noise and multi-path reflections, and are represented as 6-8  
6 bit values for the channels I and Q, as is commonly  
7 understood in digital signal processing. Figure 11b shows  
8 the result of correlating the incoming symbol stream against  
9 the Barker codewords 01001000111 and 10110111000, for the I  
10 channel of the sliding correlator 250 of figure 10. A  
11 single position in the 11 bit sequence causes a positive  
12 peak for a 1, and a negative peak for a 0, thereby  
13 establishing a synchronization window and the value of the  
14 codeword. Since the incoming baseband RF is typically phase  
15 rotated to produce an output in the I channel, the Q channel  
16 shown in figure 11c is silent. In order to satisfy the  
17 Nyquist rate for sampling, the actual sampling speed is set  
18 to 22Mhz, which is twice the bit rate of the symbols at  
19 11Mhz. Results of sampling at this rate are shown in  
20 figures 11d and 11e in the absence of multi-path  
21 reflections. For a 1Mb wireless data rate, the Barker  
22 codewords are carried only in the I channel, and for a 2Mb  
23 data rate, the 11 bit codewords are carried simultaneously  
24 in the I and Q channels. The effect of multi-path  
25 reflections is phase rotation of the analytic baseband

1 signal, so that some of the reflected energy from the I  
2 channel codewords may appear at a different time, and  
3 partially rotated into a different channel, as shown in  
4 figures 11f and 11g. The larger amplitude reflection occurs  
5 in the same position as the pure reflection shown in figures  
6 11d and 11e, and a second reflection appears in the I and Q  
7 channels shown in figures 11f and 11g. In terms relative to  
8 the largest amplitude signal within a window, signals which  
9 are part of the codeword that appear before the largest  
10 amplitude signal are called pre-cursors, and reflections  
11 which occur later are called post-cursors, as was described  
12 for figure 1.

13 Figures 12a and 12b show the 22Mhz sampled I and Q  
14 channels in the presence of a single multi-path reflection,  
15 whereby the post-cursor occurs near the middle of the  
16 symbol window and appears in the I and Q channels. Figure  
17 12d shows the training decision output 256 of figure 10,  
18 which is centered within a window that includes a provision  
19 for pre-cursors and post-cursors. Since most of the signals  
20 which follow the strongest signal are post-cursors, it is  
21 acceptable to define the window to include 5-10% pre-  
22 cursors, and 90-95% post-cursors from the largest sliding  
23 correlator output, as shown in figure 12a.

24 Figure 13a shows the contents of the channel profile  
25 memory 220 from figure 10 during the training interval. In



1 figure 10, the channel profile memory 220 comprises memory  
2 223 and memory controller 221. During the initial part of  
3 the training interval, the D(clr) output of memory  
4 controller 221 is asserted to the Din of memory 223. \_At the  
5 same time, the read/write (R/W) signal is asserted as the  
6 write and addresses (A) are cycled, thereby writing a 0  
7 value into all locations. In this manner, the contents of  
8 memory 223 are initialized at the beginning of the training  
9 sequence using the learn/decide signal 262. The channel  
10 profile memory 220 comprises a number of locations equal to  
11 the number of samples in a codeword window. For a 22Mhz  
12 sampling rate and 11 bit symbols, the channel profile memory  
13 220 comprises 22 locations for each I and Q channel. The  
14 channel profile memory may have from 2 to 20 samples of data  
15 where 6 bit data is collected and accumulated. The width of  
16 each I and Q channel location depends on the number of  
17 samples and size of each sample to be added. In general,  
18 the memory 223 width is  $d + \log_2(n)$ , where n is the number  
19 of averages, or number of times the training values are  
20 added in, and d is the intrinsic sample resolution of the  
21 correlation peaks 258. The contents of the memory 223 may  
22 be truncated to a lower number of bits for economy.  
23 Initially, the contents of the channel profile memory 223 is  
24 cleared as described earlier, and once the training decision  
25 block determines a suitable window, the channel profile

1 memory controller 221 synchronizes the storage of  
2 correlation peaks 258 with this window. When the  
3 correlation peaks 258 contain a codeword that the training  
4 decision indicates is the Barker 1 codeword 10110111000,  
5 this correlation sequence is added directly to the channel  
6 profile memory 220 as a complex value. When the Barker 0  
7 codeword of 01001000111 is received, this value is inverted  
8 and added to the channel profile memory 220. This is done  
9 for perhaps 20 cycles synchronized with each window, each  
10 time adding the current correlation peaks 258 to the  
11 previous correlation peaks already present in memory 223.  
12 Each additional cycle produces a growing value for the I  
13 channel of the channel profile memory 220 as shown in figure  
14 13a and for the Q channel of the channel profile memory 220  
15 as shown in figure 13b. In this manner, the channel profile  
16 memory 220 in figure 10 contains the correlation peaks  
17 associated with the average channel response over the  
18 training interval, and each new correlation peak is  
19 synchronized and added to the previous correlation peak.

20 After the training interval, the demultiplexer 260 in  
21 figure 10 delivers no additional codewords to the channel  
22 profile memory, and the memory controller 221 prevents  
23 additional data from being written using learn/decide signal  
24 262. The contents of channel profile memory 223 are  
25 sequentially read by memory controller 221 in synchronism

1 with the incoming correlation peaks using window signal 288.  
2 The contents of channel profile memory 223 which are now  
3 synchronized with the incoming correlation peaks 258, are  
4 conjugated by complex conjugator 272. This conjugation  
5 nullifies the channel effects due to varying phase of the  
6 reflected signals. This conjugation changes the sign of the  
7 imaginary part of the complex contents of memory 223. The  
8 output of the complex conjugator 272 is multiplied with the  
9 correlation peaks 258 within complex multiplier 278. These  
10 aforementioned correlation peaks 258 are selected by the  
11 demultiplexer 260 using the learn/decide signal 262, and the  
12 multiplier 278 output is delivered to an accumulator 294.  
13 The memory 289 within the accumulator 294 is initialized at  
14 the beginning of the window signal 288, and incoming  
15 multiplier output values are summed with the present value  
16 of memory 289 using adder 287. At the end of the window as  
17 indicated by signal 288, the values of the I and Q channels  
18 of the accumulator are examined. Since the pre-cursors and  
19 post-cursors have been added into these resulting values for  
20 I and Q, the resulting contents of memory 289 is then  
21 compared 291 with a threshold(0) to determine if the value  
22 is 1 or 0 in each of the I and Q channels. In this manner,  
23 the accumulations of all pre-cursors and post-cursors in the  
24 window belonging to a particular codeword are provided to  
25 the I and Q channel outputs, and a decision is made on these

1 symbols. The ability to make a decision on these symbols  
2 generated in the described manner has reduced complexity and  
3 increased performance compared to prior art solutions.

4 Comparing the functionality and complexity of the  
5 combiner of figure 10 to the prior art of figures 5 or 6, a  
6 clear reduction in complexity has been achieved with no loss  
7 in functionality. The various delayed signals enter figure  
8 5 input 194, and a plurality of separate delays and  
9 multiplication coefficients are calculated, conjugated,  
10 multiplied, and summed to the final result. A full  
11 implementation of the signal processor 190 on a 22 sample  
12 per symbol incoming rate would require 22 such channels,  
13 each with a complex multiplier. Typically, a much smaller  
14 number of channels are implemented, such as the 3 shown in  
15 figures 5 & 6, and these channels operate on a few large  
16 signals which are selected for phase correction. The  
17 decision 188 is made on this small number of signals. In  
18 the present invention shown in figure 10, all of the signal  
19 reflections are contained in the channel profile memory 220  
20 of figure 10, and each sample interval, a current channel  
21 memory content value is conjugated, multiplied by incoming  
22 correlation peaks 258, and summed in accumulator 294. In  
23 this manner, every sample is used to produce the final  
24 accumulated result stored in memory 289, in contrast with  
25 the finite number of delay paths associated with figures 5

1 or 6. For a system of figure 5 or 6 to achieve this  
2 performance at 22 samples per symbol, 22 delay elements (170  
3 and 172) and 22 multipliers (180, 182, and 184) would be  
4 required. The system of figure 10 achieves this same  
5 functionality with a single multiplier 278. The reference  
6 to the present invention as an "all-tap" combiner makes  
7 reference to the equivalent functionality of a 22 tap FIR  
8 filter 203 of figure 6 being achieved in the single  
9 multiplier circuit of figure 10.

10 An additional advantage of the present invention  
11 compared to the prior art of figures 5 or 6 is simplified  
12 circuitry allows operation at twice the incoming sample  
13 rate. The incoming 11Mbps data stream may be sampled at  
14 22Mhz as presented in the data stream 286 of figure 10. The  
15 peaks of the correlator outputs in 1us duration are used to  
16 create the channel profile, which is averaged over many  
17 cycles to reduce the noise content. When the channel  
18 profile memory contents are multiplied by the incoming data  
19 stream at 22Mhz, the phase offsets associated with initial  
20 timing offsets are cancelled. If the rake training is  
21 continued throughout the packet such that the channel memory  
22 220 continues to add new correlation peak data 258 over the  
23 duration of packet reception, as opposed to stopping the  
24 learning process after the preamble, the need for a timing-  
25 tracking loop (not shown) to track drifts between the

1 transmitter oscillator (not shown) and received data clock  
2 is eliminated, as the slow phase change of the incoming data  
3 is compensated by the slow phase change of the contents of  
4 the channel profile memory.

5 Figure 10a shows the alternate functionality, where the  
6 incoming correlation peaks 258 are delivered to multiplexer  
7 280a, which always sends the peaks to channel profile memory  
8 264, and the memory controller 221 of figure 10 is always  
9 adding incoming data to the channel profile memory 223,  
10 regardless of the state of learn/decide signal 262, which  
11 now controls switch 261, enabling data to the complex  
12 multiplier 278 when the learn/decide signal 262 is in the  
13 decide state. As both the channel profile memory and  
14 incoming data slowly rotate in phase together because of  
15 slow drifts in clock, the result of conjugation and  
16 multiplication of the channel profile memory with the  
17 incoming correlation peaks is the rotation of the  
18 multiplication product back into the correct I and Q  
19 channels.

20 Figures 14a and 14b show the overall operation of the  
21 all-tap, fractionally-spaced serial rake combiner processor  
22 284 in figure 10. In figures 14a and 14b, learning interval  
23 300 is followed by a decision interval 302. During the  
24 learning interval 300, sliding correlator outputs I 314 and  
25 Q 320 are generated by sliding correlator 250 of figure 10,

1 and then as also shown in figure 10, training decision  
2 function 252 generates training decision output 256 and  
3 window signal 288. When the training decision output 256 of  
4 figure 10 which is illustrated by the waveform 304 of figure  
5 14a is 1, multiplexer 292 selects the correlation peaks 258  
6 through demultiplexer 260. When the training decision  
7 output 256 is 0, the correlation peak values are inverted  
8 with inverter 290 during the duration of the window signal  
9 288, and inverted values are selected through multiplexer  
10 292. In this manner, same-polarity symbols including multi-  
11 path reflections are successively added into the channel  
12 profile memory 220 using adder 268 of figure 10. Waveform  
13 306 of figure 14a shows the same-polarity training  
14 multiplexer output found at the multiplexer 292 of figure  
15 10. This same-polarity training multiplexer output is either  
16 the original sliding correlator output when the training  
17 decision 304 is 1, or the negated value when the training  
18 decision 304 is 0. Waveform 308 of figure 14a shows the  
19 contents of channel profile memory 220 of figure 10, which  
20 initialize to 0 at the start of the training interval, and  
21 grow with each successive addition 268 from the multiplexer  
22 292. This multiplexer 292 output is illustrated in waveform  
23 306 of figure 14a. At the end of the learning interval 300  
24 of figure 14a, the channel profile memory 220 of figure 10  
25 has the additive contribution of signals from each cycle,

1 while the random noise component tends to average out. This  
2 averaging thereby achieves an improved signal to noise  
3 version of the correlation peak to use as a template during  
4 the resulting decision interval 302 of figure 14a.

5 Figure 14b shows the contents of channel profile memory  
6 220 of figure 10 during the learning interval 300  
7 illustrated by waveform 308/322. During the learning  
8 interval 300, the window waveform 310 is used by the channel  
9 profile memory controller 221 of channel profile memory 220  
10 to establish synchronization between incoming data from  
11 multiplexer 292 and training window data stored in channel  
12 profile memory. During decision interval 302, window  
13 waveform 310 defines the boundaries of each codeword and  
14 indicates the beginning and end of each accumulator 294 of  
15 figure 10 window. The window 310 and training decision 304  
16 waveforms are asserted at the earliest point that the  
17 training decision function 252 of figure 10 is able to  
18 ascertain the largest signal to use as a datum reference, as  
19 was described earlier. The output of the complex multiplier  
20 278 of figure 10, is illustrated by waveform 322 of figure  
21 14b. Waveform 322 shows the result of multiplying the  
22 incoming sliding correlator output 314/210 by the complex  
23 conjugate of the channel profile memory waveform 308 of  
24 figure 14a. The window signal 288 of figure 10 and window  
25 waveform 310 of figure 14a have been centered by the



1 training decision function 252 of figure 10 about the  
2 strongest correlation peak to include some pre-cursors, and  
3 the majority of post-cursors. This window waveform 310 of  
4 figure 14a is used to reset the accumulator 294 of figure  
5 10 at the beginning of each window. After the multiplier  
6 output values are accumulated during the acquisition window,  
7 the value is sampled at the end of the acquisition window  
8 and a decision is made, as shown in the decision values of  
9 waveform 324 of figure 14b. This accumulation and sampling  
10 operation is performed on the I and Q channels of the  
11 accumulator 294 of figure 10.

12 As the result of the aforementioned process, a decision  
13 can be made from a stream of quadrature input signals which  
14 conform to the IEEE 802.11b wireless data rates of 1Mb and  
15 2Mb comprising Barker codes. It is clear to one skilled in  
16 the art of digital signal processing and wireless encoding  
17 systems that the steps of training a channel profile memory  
18 require the Barker code for 0 be inverted compared to the  
19 value of the Barker code for 1. Within the system described,  
20 it is possible to do this for the I and Q channels, thereby  
21 collecting an averaged Barker codeword for the channel. The  
22 system described comprises, as illustrated in figure 10, a  
23 sliding correlator 250 for the recovery of codewords which  
24 may include reflected codewords and a training decision  
25 function 252 for generation of a window signal for framing

1 the codewords. The system also includes a training decision  
2 output 256 for determining whether the codeword for 0 or 1  
3 was sent to accumulate same-polarity symbols into the  
4 channel profile memory 220, which contains a sum of  
5 responses.